RAYTHEON

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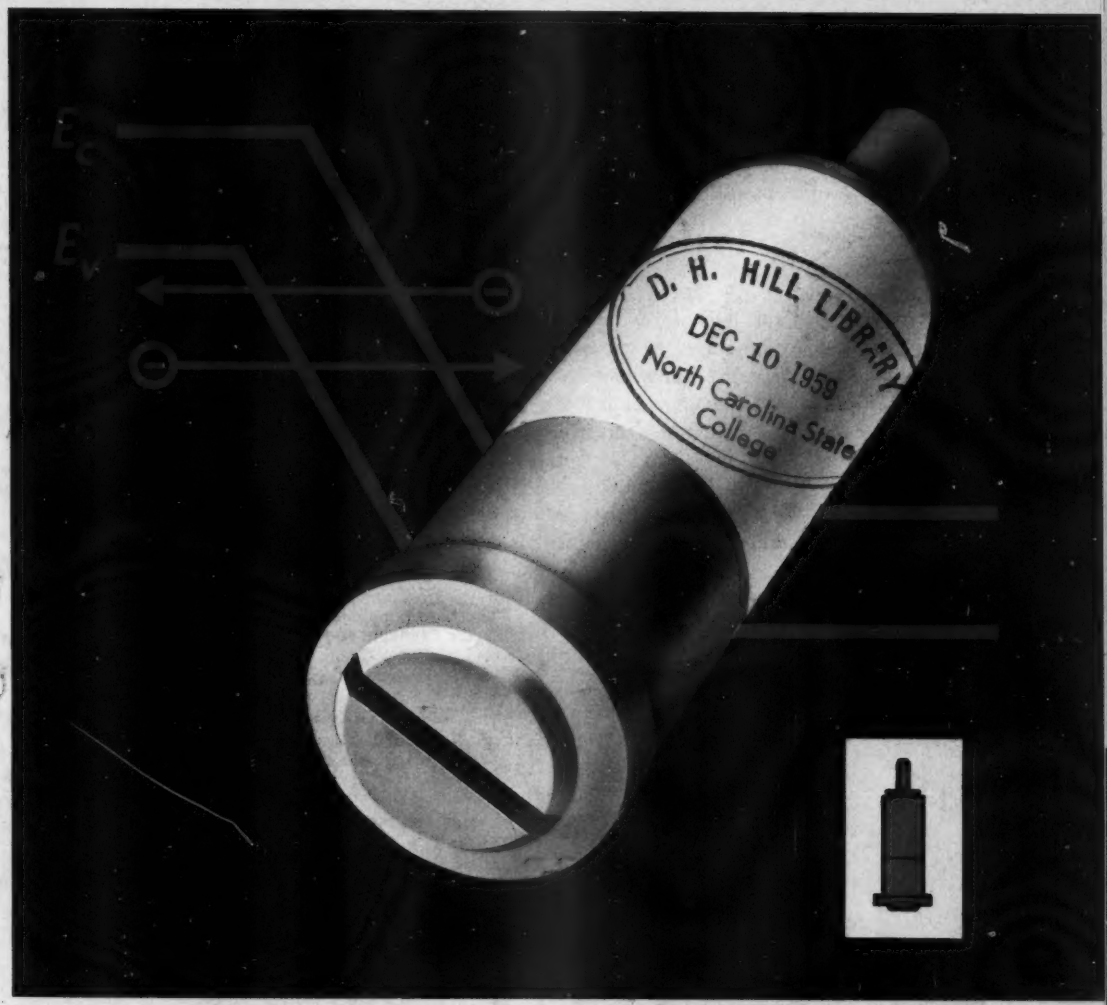

AN ESAKI "TUNNEL DIODE"

In this issue: LOW NOISE MICROWAVE AMPLIFIERS

ELECTRONIC PROGRESS

Volume IV November-December 1959 Number 3

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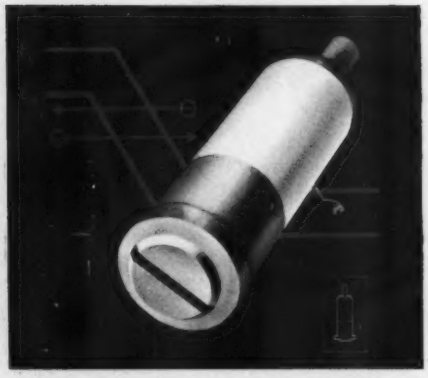

One of the newest developments in the low noise microwave amplifier field is the Esaki "tunnel diode." The diagram shown in the background is used to represent the "tunneling" of carriers across an energy gap, the basic mechanism upon which this new semiconductor device depends for amplification.

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Calendar

ENGINEERING SEMINARS

Seminars will be held every other Tuesday at 4:45 P.M. in various laboratory cafeterias. Security clearance will be required as indicated. Personnel who do not hold picture badges should consult their security officers for attendance arrangements.

"Objectives and Management of Engineering and Research at Raytheon"	Nov. 3	Confidential
Ivan A. Getting, Vice President, Engineering and Research	Spencer Lab Cafeteria	
"The FPS-28 Radar"	Nov. 17	Confidential
W. L. Pritchard, Director of Engineering Heavy Electronic Equipment Subdivision	Bedford Systems Lab Cafeteria	
"The Government Equipment and Systems Group —Its Products and Plans"	Dec. 1	Confidential
D. D. Coffin, Group Vice President, Government Equipment and Systems Divisions	Wayland Lab Cafeteria	
"The Machlett Division"	Dec. 15	Unclassified
J. W. Skehan, Vice President in charge of Manufacturing and Engineering, Machlett Laboratories, Inc.	Spencer Lab Cafeteria	
"Design and Standardization for Weapon Systems"	Jan. 5	Unclassified
J. R. Townsend, Special Assistant to the Director of Defense Research and Engineering, Department of Defense	Spencer Lab Cafeteria	

Contents

Low Noise Microwave Amplifiers ..... *L. Davis, Jr.* 1
Publications and Papers 5
The Maser ..... *H. Statz* 6
The Parametric Amplifier ..... *S. Cohen* 11
The Esaki "Tunnel Diode" ..... *R. Pucel* 16
Technical Notebooks 20

LOW NOISE MICROWAVE AMPLIFIERS

L. Davis, Jr.
Research Division

THERE has recently been a great deal of interest, and perhaps an undue amount of excitement, about a variety of solid state low noise amplifiers capable of operation in the microwave frequency region. Quite naturally there has been a great deal of interest here at Raytheon in such amplifiers because of the large fraction of our business in microwave systems. These amplifiers are classified generally as masers, parametric amplifiers, and Esaki tunnel diode amplifiers. They are most commonly one-port negative-resistance devices amplifying the microwave frequency power impinging on the input terminals. Hence they would, in general, be used as preamplifiers before conventional amplifiers or crystal mixers in order to lower the noise figure of the over-all receiver. In this article we want to review the actual achievements of these new solid state amplifiers in terms of noise figures and frequencies of operation, and to compare them with conventional amplifiers. Following this we consider various types of microwave systems and discuss the factors involved in estimating the system improvement that these amplifiers make possible. Finally we discuss some applications for which low noise amplifiers are particularly valuable.

In discussing low noise amplifiers, particularly very low noise amplifiers such as masers, it is well to keep in mind the definition of noise figure and also the concept of noise temperature. The noise figure of a system is defined as the signal-to-noise ratio at the antenna, divided by the signal-to-noise ratio at the output of the amplifier and is usually expressed in decibels. It is general practice to define the noise at the antenna as that due to Johnson noise generated by a 300°K matched source. This is approximately the case for an antenna looking predominantly at the surface of the earth. The noise temperature of a device is defined in terms of the excess noise observed at the output of the device. It is equal to the tempera-

Recent developments in the use of solid state devices for low noise amplification have made possible increasingly sensitive detection of small signals. Because of the obvious significance of these developments to the fields of radar and communications, we have devoted an entire issue to the topic of low noise microwave amplifiers. A general article is followed by detailed treatments of the three main types: the maser, the Esaki tunnel diode, and the parametric amplifier.

ture of a matched input resistor that would lead to this same output noise in the absence of noise generated within the device. Thus, for example, when one says that a receiver has a 6 db noise figure, one means that the noise found in the output of the receiver is four times the noise that would be present in a perfect receiver if the antenna were looking at a 300°K black body. The noise temperature of the receiver corresponding to this excess noise is three times 300°K, or 900°K.

Masers have been built to operate at frequencies from 400 megacycles (used in the Venus echo experiment at Lincoln Laboratory) up through X-band, and can certainly be built for much higher frequencies. Suitable paramagnetic maser materials still must be discovered for use at these higher frequencies, but this is a promising and rapidly developing field of research. A real limitation is the need for sources of microwave power, of the order of several milliwatts, at a frequency necessarily higher than the signal frequency. Maser amplifiers have been built with noise temperatures (of the amplifier alone) of 2°K. For such low noise temperatures they must be operated at liquid helium temperatures, but it has been found possible to operate them successfully at temperatures as high as that of liquid nitrogen under certain conditions, with consequently higher noise temperatures.

Parametric amplifiers have, in general, higher noise temperatures than masers. In the L-band region they have been built with noise temperatures as low as 70°K. This means that in conventional receivers a parametric amplifier operating as a preamplifier can be used to reduce the noise figure to values as low as 1 db. The parametric amplifier, like the maser, must have a source of rf pump power of frequency higher than that of the signal to be amplified, and the noise performance improves as the ratio of pump frequency to signal frequency. Up to now parametric amplifiers have been limited to frequencies of the order of 10 kmc but an extension to higher frequencies will surely take place as suitable diodes are developed. In general, noise figures of the order of 3 db are at present to be expected for the higher frequencies.

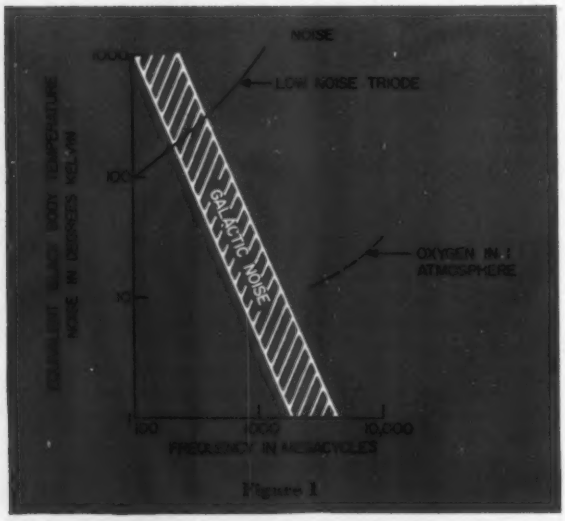


Figure 1

Esaki diodes are characterized by very large capacitance and hence very low impedance at microwave frequencies. At frequencies of S to X-band they may exhibit impedances of about one-tenth of an ohm or less and therefore present some difficulties in circuit techniques at microwave frequencies. Aside from the disadvantage of very low impedance, Esaki diodes have no inherent frequency limitation up to frequencies of the order of 100 kmc. They derive their energy from a dc supply and thus do not require an rf pump. This is a major advantage, particularly at extremely high frequencies. The chief disadvantage is that the Esaki diode is inherently more noisy than the parametric amplifier since it has shot noise in addition to Johnson noise. The lowest noise temperature achieved to date is about 200°K or a little better than a 2 db noise figure in a conventional receiver. Most of the Esaki diodes reported have operated at quite low frequencies, but there is a strong development effort in this area which should lead to high frequency diodes in the near future.

How do these noise figures compare with those of presently used receivers? In the low frequency region, of the order of a few hundred mc, triodes are available with 1 or 2 db noise figures. Traveling wave tubes are reported with noise figures at L-band as low as 4 db, and as low as 4.5 even at X-band. Crystal mixers have noise figures as low as 6 db at S-band rising to the 12 or 15 db range at the millimeter end of the spectrum. One must consider the noise figure of presently available receivers before trying to consider the use of one of the new low noise solid state amplifiers. It is, for example, of little use to use an Esaki diode preamplifier before a very low noise amplifier tube at, say, 400 megacycles.

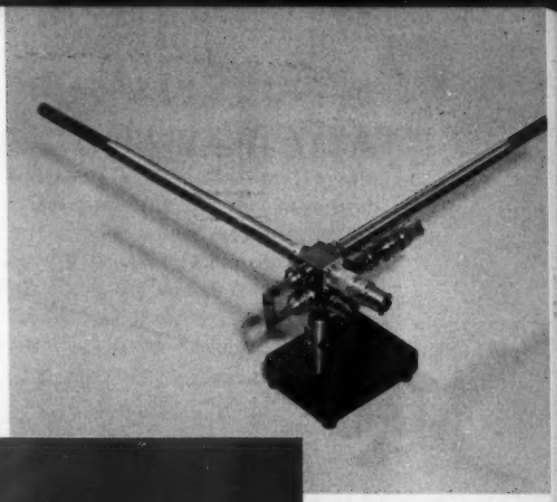
To obtain some measure of the importance of a low noise amplifier to a specific system, the definition of noise figure (or noise temperature) given above must be applied to the over-all receiver. One must consider very carefully all sources of noise from the antenna input, through the various attenuation losses and stages of amplification and mixing, to the receiver output. Only when one compares complete receivers, with and without low-noise amplifiers, can one accurately estimate the value of the low noise amplifier.

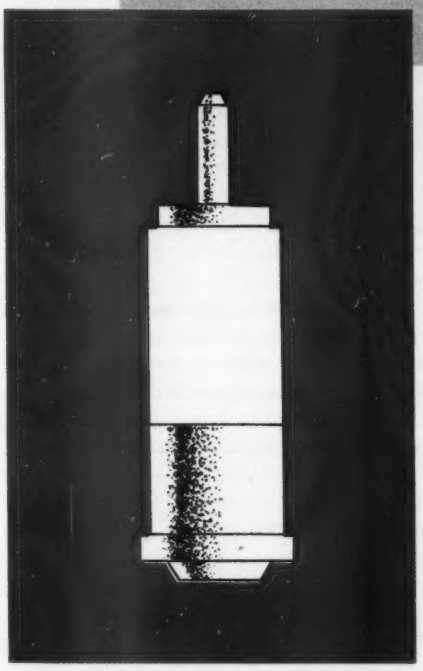
The first factor to be considered is, naturally enough, the effective noise at the antenna input. Just as there is little sense in improving the noise figure of a con-

ventional broadcast radio receiver in the presence of radio frequency interference from automobile ignition, there is little sense in putting a 2°K noise temperature maser in a radar system that looks at the surface of the earth at 300°K. Search radars that search the horizon, and hence look at the surface of the earth, or airborne radars that map the surface of the earth, obviously fall in this class and have antenna noise temperatures close to 300°K. But what noise temperature does the antenna see if it looks somewhere in the sky? This is a little more complicated because it depends first of all on the details of the antenna radiation pattern, that is, the amplitude and position of the side lobes. If the side lobes see the surface of the earth, then, of course, they will receive a certain amount of noise from this 300°K noise source, and the total integrated side lobe noise reception from the surface of the earth must be considered. This factor will clearly vary greatly from system to system.

In addition, we know that if an antenna looks at the sun, it sees a very noisy source. This is usually, at worst, a transient condition which is not significant for the type of receiver considered here. However, there are other sources of noise in the sky, both within our own galaxy and also in the further reaches of the universe. Several mechanisms such as thermal noise from stars and synchroton radiation from interstellar space contribute to this noise. Our own atmosphere produces attenuation of microwave energy by means of various molecular absorption bands and tends to become more opaque at high microwave frequencies. It thus contributes a noise corresponding to its temperature and proportional to its absorption at the frequency of operation. These two sources of noise are indicated in Figure 1. It will be noticed that the galactic noise covers quite a spread in noise temperature. This is primarily due to the fact that it is position-dependent, some portions of the sky being much "hotter" than others. These various sources of noise must now be summed to obtain the noise temperature at the input of the antenna, the base for the noise figure of the over-all receiver.

In many applications we must also consider interference from both friend and enemy. This question is difficult to treat in a very general way because of the great variety of tactical usages to be considered. For example, a fire control system which has been built to have necessary range performance in the face of given types of enemy jamming probably has more than adequate range performance when not faced with jamming. Therefore, the addition of a low noise receiver is probably not of much importance to such a system. However, where the additional price to be





Parametric amplifier
Esaki tunnel diode
Maser

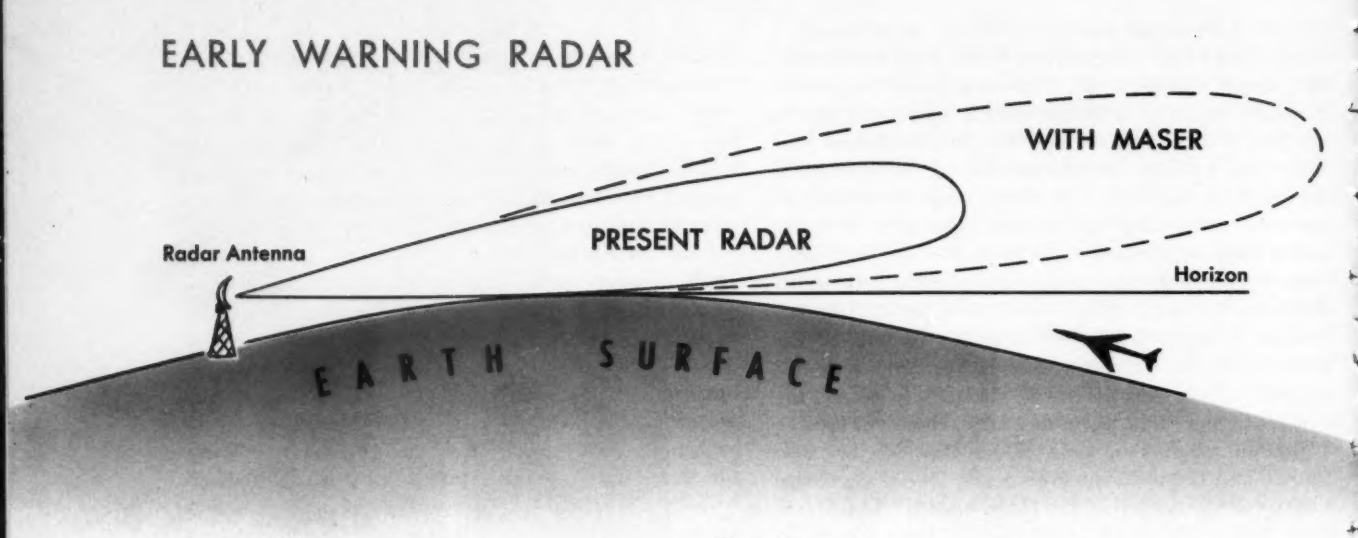


Figure 2

paid is low, as would be the case for a parametric amplifier, then there is perhaps some justification for incorporating a low noise preamplifier in order to give a little longer range search capability to the fire control system. A surface-based extremely long range early warning radar for observation of aerodynamic vehicles is severely horizon limited (see Figure 2). In such a case there is very little to be gained by the addition of low noise amplifiers. Such radars usually have sufficient transmitter power to achieve the maximum range allowed by the horizon of the earth against aerodynamic vehicles in the presence of enemy jamming. This means that without enemy jamming any improvement that can be achieved with a low noise amplifier would really be negligible in terms of excess range picked up beyond the horizon.

One may now ask, what is the price one has to pay for these amplifiers? The maser, by far the most complex, requires cryogenic techniques and extreme care in the over-all circuitry of the system to utilize its very low noise temperature. Because the attenuation of an appreciable length of wave guide gives rise to a significant amount of noise, the maser should be placed directly at the focus of the antenna if possible. This is particularly inconvenient because of the presence of the cryostat and the need to preserve the cryogenic fluid (usually liquid helium) while the antenna is scanning. As mentioned previously, one ordinarily prefers to use the maser, as well as the other microwave low noise amplifiers, as a one-port device. With a circulator at the input the maser does not see the noise generated in the next stage of the receiver, except for a small amount due to finite isolation and to load mismatch. The forward loss of the circulator should be extremely small since every tenth of a decibel contributes about 7°K of noise. The maser, finally, requires a permanent magnet and a microwave pump source at a frequency higher than the signal frequency. All this adds up to considerable equipment.

The parametric amplifier, on the other hand, does not require either the magnetic field or the cryostat. It does require a microwave pump, and this pump should be extremely stable if gain or phase stability is required. The Esaki diode is least complicated in that it does not require a microwave pump. Instead it merely uses a dc bias supply for its source of energy. In the cases of both the parametric amplifier and the Esaki diode a microwave circulator with insertion loss of less than several tenths of a decibel should be used.

Where can we expect to find these amplifiers used? Probably masers will not find much usage except in space applications and perhaps, sometime in the future, in passive microwave detection systems. It is only if the antenna looks at space that really low temperature inputs at the antenna are realized. For an antenna on the earth's surface searching for an object flying in space we should use a frequency somewhere between 3 and 10 kmc to avoid high noise temperature at the antenna (shown by Figure 1). In this frequency range the use of a maser certainly looks justified. For satellite-to-satellite communications or for satellite-based radar, again a maser can in all probability be justified in spite of its requirement of cryogenic techniques.

For passive microwave mapping, bandwidth of the receiving system must be considered along with noise figure. The noise temperature at the antenna is of the order of 300°K so that we gain very little in

getting noise figures much below 1 db. The temperature differences that can be measured are in inverse proportion to the square root of the bandwidth, so that the bandwidth of the system becomes the important factor once we get a reasonably low noise figure. To get improvement over present traveling wave tube techniques we must have very wide band masers or parametric amplifiers. Both masers and parametric amplifiers of the single-cavity type give fractional voltage gain-bandwidth products not much higher than 1 to 2 per cent. It therefore would seem that a clear advantage over traveling wave tubes will require development of traveling wave masers or parametric amplifiers which will give voltage gain-bandwidth products of the order of 50 per cent of the operating frequency. Since the low noise temperature of the maser is not important in this application, the traveling wave parametric amplifier will probably be preferred because of its greater simplicity.

Esaki diodes seem to show promise for relatively low noise amplification at those frequencies at which we cannot hope to achieve parametric amplification, either because of diode or pump limitations. A parametric amplifier operating at 60 kmc would require diodes that are not available now and that have properties approaching the best values theoretically possible. Furthermore, for good noise figures the pump should operate well above the signal frequency with something like 10 milliwatts of power. This too seems to be a serious limitation. In this frequency range the Esaki diode amplifier offers considerable improvement since present amplifiers have noise figures of the order of 15 db. The chief problem will be in learning the circuitry and the device techniques to overcome the extremely low impedances found at these frequencies.

In summary, we should say that although we will find some very important uses in the microwave frequency range for masers and for Esaki diode amplifiers, the parametric amplifier will in all probability find the broadest range of applications because of its flexibility, low noise, and relative simplicity of equipment. However, each type of amplifier does offer its own advantages and these should be weighed carefully when deciding which is most applicable to a given system.

Publications & Papers

H. R. BULLOCK, "The Design of Light Weight Casting," presented at the SAMPE Meeting in New York on May 15, 1959.

C. KLEIN, W. STRAUB, "Radiation-Induced Energy Levels in Silicon: The Present State of Affairs," presented at the Conference on Radiation Effects in Semiconductors in Oak Ridge on May 6, 1959.

C. KLEIN, "Radiation-Induced Energy Levels in Silicon," published in the JOURNAL OF APPLIED PHYSICS, August, 1959.

R. A. LARSEN, JR., L. M. WINSLOW, "Video Mapping Inverter," published in SPACE AGE, July, 1959.

J. LAVINE, "The Ordinary Hall Effect in Fe_3O_4 and $(NiO)_{0.75}$ $(FeO)_{0.25}$ at Room Temperatures," published in PHYSICAL REVIEW, April, 1959.

A. L. MacDONALD, W. J. HEINECKE, "Effects of Heat Treatment of Diffused Silicon Material," presented at the Electrochemical Society Meeting in Philadelphia on May 4, 1959.

D. McMURTRIE, "Magnetic Amplifier Circuits — A Classification of Half-Wave and Full-Wave Non-Reversible Self-Saturating Circuits," presented at the AIEE Summer Meeting in Seattle on June 26, 1959.

H. L. MYERS, "Meetings, Metals and Methods," published in AMERICAN MACHINIST, June, 1959.

A. E. O'BRIEN, E. RICHTER, "Tunable Spherical Cavity Resonators," presented at the meeting of the IRE (PGMTT) in Cambridge on June 2, 1959.

J. OSEPCHUK, "Sorting in M-type Traveling-Wave Tubes," presented at the Conference on Electron Tube Research in Mexico City on June 25, 1959.

A. E. PALADINO, "Phase Equilibria in the Ferrite Region of the System Fe-Mg-O," presented at the American Ceramic Society Meeting in Chicago on May 19, 1959.

G. PARTRIDGE, "A Transistorized Pulse Code Repeater," presented at the AIEE Summer Meeting in Seattle on June 25, 1959.

F. A. PLEMENOS, "Terrain Bias Error in Doppler Navigational Radars," presented at the Symposium on Radar Return in Albuquerque on May 12, 1959.

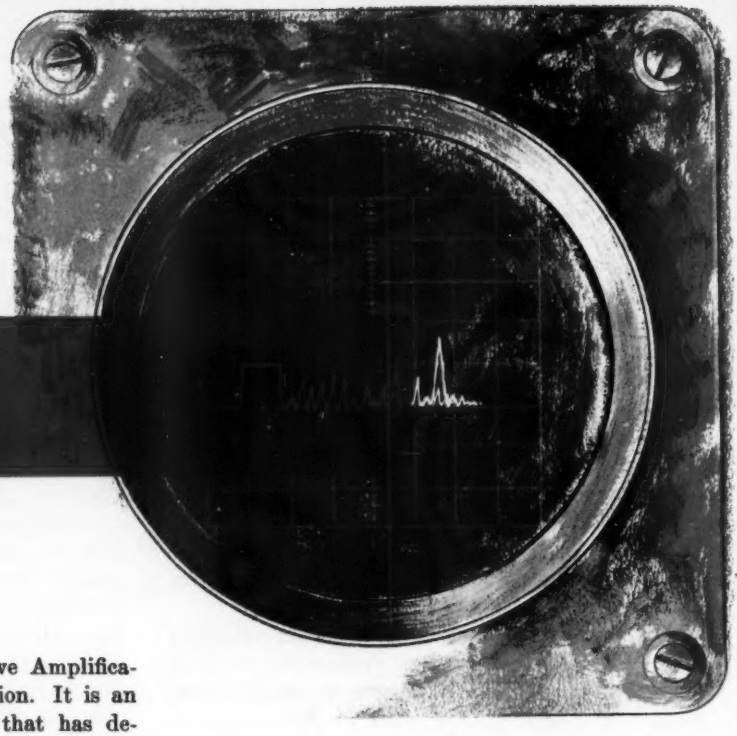
D. SCHIFF, "Bonding Experiments with Exploding Foils," presented at the AFCRC Conference on Exploding Wires in Boston on April 2, 1959.

B. H. SMITH, "A Monotron Oscillator Amplifier Device," presented at the Conference on Electron Tube Research in Mexico City on June 26, 1959.

T. C. TAYLOR, "Fused Contacts to Silicon Carbide," presented at the AFCRC Conference on Silicon Carbide in Boston on April 3, 1959.

H. Statz
Research Division

the MASER



Introduction

The word maser stands for Microwave Amplification by Stimulated Emission of Radiation. It is an entirely new concept of amplification that has developed since World War II. In the following article we want to explain the principles of operation of the maser together with its possible applications and its potential of leading to other new and unconventional devices.

In the maser use is made of individual free atoms or ions built into crystal lattices to produce the functions of an amplifier or an oscillator. In a certain sense the atoms and ions may be considered themselves as very tiny oscillators with natural frequencies proportional to the separations between pairs of energy levels. The natural frequencies of these tiny oscillators are usually quite high, and the wavelengths of the radiation produced by them may reach from the infrared to the ultraviolet and to the X-ray region. If we want to use these oscillators in the conventional microwave region (wavelengths of a few centimeters), then most of the transitions are of too high a frequency and some rather special transitions have to be considered.

Frequencies in the microwave range are obtained by letting an electronic spin precess in a magnetic field of a few thousand gauss. Also, some vibrations of molecules show frequencies in the microwave spectrum. The first maser was a so-called "gas" maser. At present the emphasis of the research and development is on the solid state maser which was suggested in its most practical form by Professor Bloembergen of Harvard in 1956. This maser has the advantage that its frequency can be changed by adjusting a magnetic field. In this maser paramagnetic ions precessing in a magnetic field represent the essential atomic oscillators.

The maser was received with great interest since it is inherently a very low noise amplifier. As a matter of fact it is such a low noise device that it is at present entirely limited in its performance by the losses in external circuit components. Were it not for these losses it should be possible to detect the thermal or black body radiation of an object at a temperature of a few degrees Kelvin.

Principle of Operation of the Solid State Maser

One of the most commonly used materials is ruby, that is, alumina containing a small concentration of chromium. The properties to be discussed are due entirely to the paramagnetic chromium ions, while the alumina host material acts mainly as a diluting agent to decrease the strength of the interactions among these active centers. In the triply charged chromium ion one of the electronic shells, the third shell, is incompletely filled and the free ion thus has a net angular momentum and a net magnetic moment. In spectroscopic notation the lowest state is a $^4F_{3/2}$. When this ion is put into the crystal and when simultaneously a magnetic field is applied, this lowest state splits up into four distinct states separated in energy by an amount depending on the magnitude of the magnetic field and on the orientation of the magnetic field with respect to the crystalline axes. For the proper design of a maser it is important to be able to calculate these energy levels accurately. Up to date, an "effective spin Hamiltonian" formalism has most commonly been used. For example, the energy levels

for ruby are given by the eigenvalues of the quantum mechanical Hamiltonian, H_{eff}, which is of the form:

$$H_{eff} = DS_z^2 + \beta \left[g_{\parallel} S_z H_z + g_{\perp} (S_x H_x + S_y H_y) \right].$$

This Hamiltonian acts between the four substates of a spin $S = 3/2$. D is a constant related to the energy level splitting in zero magnetic field. β is the Bohr magneton, and $g_{\parallel}$ and $g_{\perp}$ are the components of the anisotropic spectroscopic splitting factor. These empirical constants describe the interaction of the ion with the lattice and the magnetic field; H_x, H_y and H_z are the components of the magnetic field with respect to the crystalline axes. Without going into mathematical details we show in Figure 1 the result of the computation of the energy levels as a function of magnetic field. Energies have been directly converted into frequencies by the use of the relation $E = h\upsilon$. In this figure the magnetic field has been chosen to make an angle of 90° with the crystalline axis (that is, $H_z = 0$); a similar figure could be derived for any other desired orientation.

For a solid state maser it is, in general, necessary to have at least three energy levels, even though there exist also two-level masers for pulsed operation only.

In Figure 2 we have depicted three of the four energy levels of Figure 1 to explain the maser action.

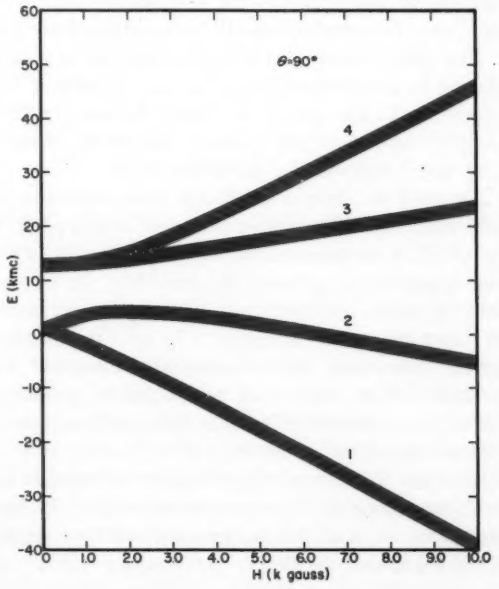


Figure 1

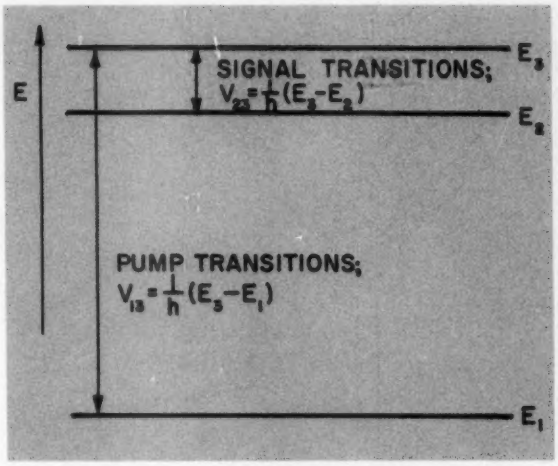


Figure 2

Without any rf fields the occupation of the levels is given by a Boltzmann distribution; that is,

$$n_i = C \exp\left(-\frac{E_i}{kT}\right) \qquad (2)$$

where n_i is the occupation of the i-th level, E_i is its energy, k is Boltzmann's constant, and T is the absolute temperature. Equation (2) states that the lower levels are more heavily occupied. In the operation of a maser a strong rf field, the so-called pump field, is applied at a frequency $\upsilon = h^{-1}(E_3 - E_1)$. This rf field causes transitions from level 1 to level 3 with absorption of energy from the field and also induces transitions from level 3 to level 1 with emission of energy. These transition probabilities can be shown to be equal for the up and for the down transitions so that with $n_1 > n_3$ there is always a net absorption of rf energy. If the rf field is now made strong enough so that the thermal relaxation mechanisms cannot compete with the pump field in establishing a Boltzmann distribution between levels 1 and 3, it follows that the occupation of these two levels must approach equality and the transition is said to be saturated.

Let us next consider the occupation of level 2. There are now two competing thermal relaxation processes. On the one hand, level 2 would like to be less heavily populated than level 1, as it would be in a normal Boltzmann distribution. By the same argument, however, 2 would like to be more heavily populated than 3. Since 1 and 3 are equally populated, this leads to a contradiction. To take into account the competing effects of the rf field and the thermal relaxation processes, one really requires a knowledge of the rate at which the thermal equilibrium distribution is estab-

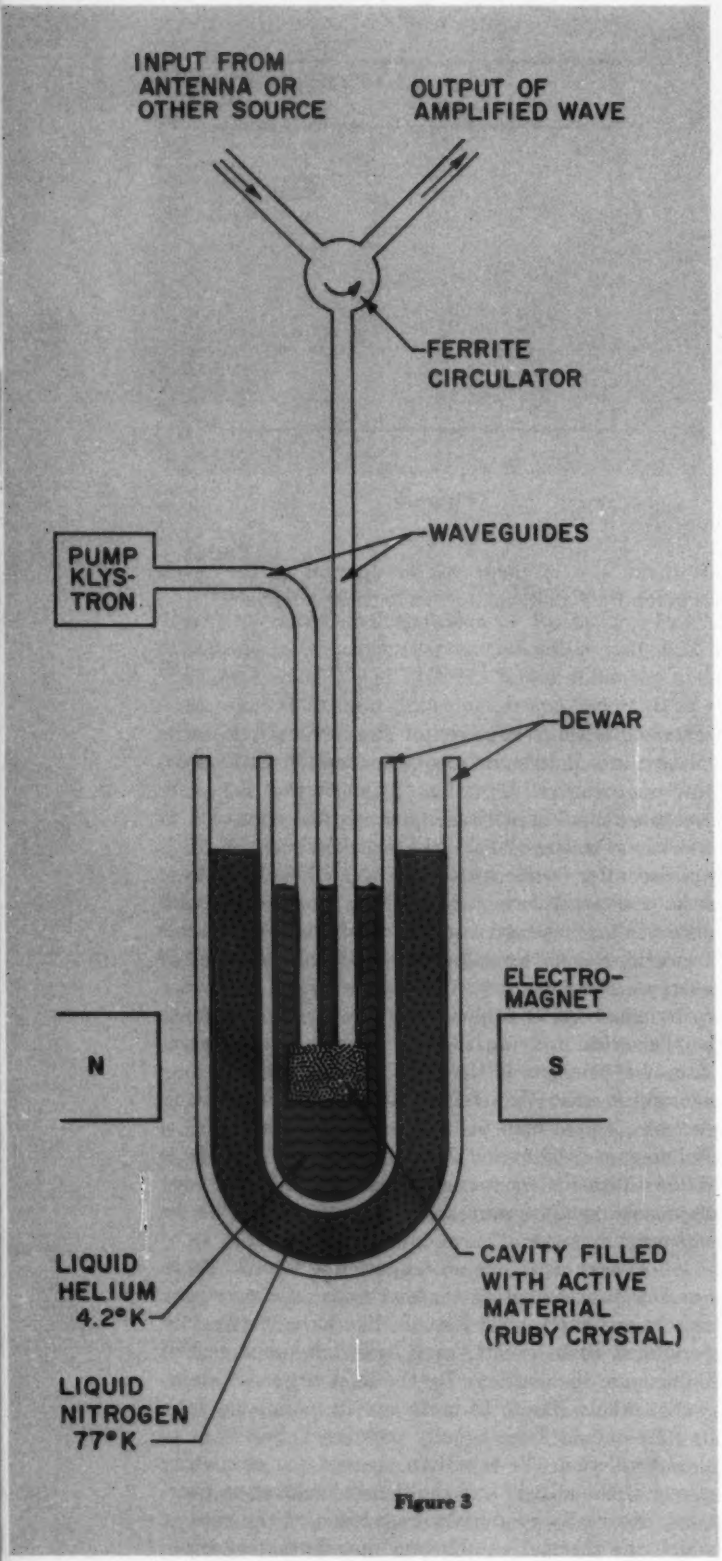


Figure 3

lished between each pair of levels and this information is not generally available. In many cases, the thermal transition probabilities will be nearly equal, and with this assumption, it can be shown rather simply that:

$$n_3 - n_2 = \frac{N}{8}\frac{h}{kT}(\upsilon_{12} - \upsilon_{23}) \qquad (3)$$

where N is the total number of atoms per unit volume, and υ_{12}, for example, stands for $h^{-1}(E_2 - E_1)$. From Equation (3) we see that level 3 is more heavily populated than 2 when the separation between levels 1 and 2 is larger than the separation between levels 2 and 3. The existence of such an inverted population is the essential ingredient of all maser-type devices. From Equation (2), it follows that an inverted population corresponds formally to a negative value of T, and thus a maser is sometimes referred to as a device utilizing "negative absolute temperatures." It is to be noted that achieving an inverted distribution requires pumping at a frequency greater than twice the signal frequency.

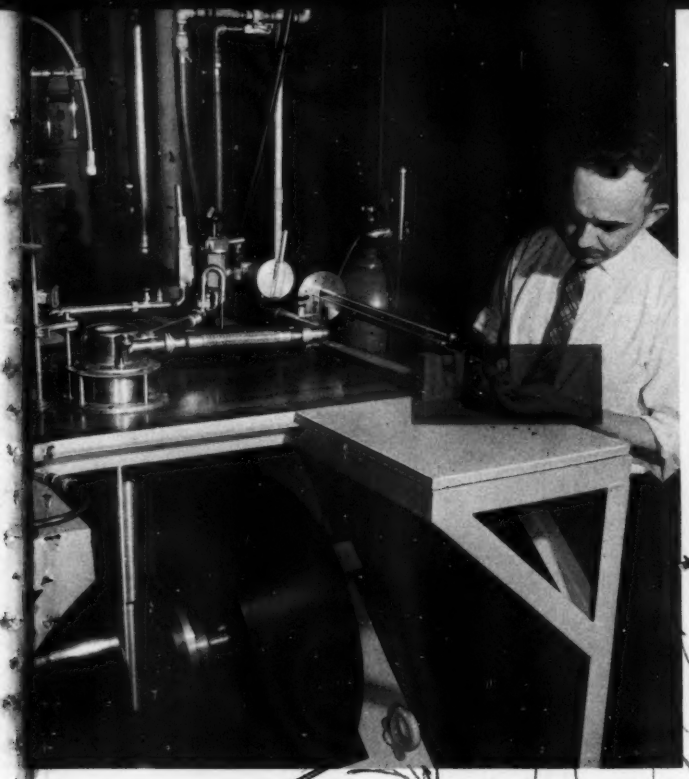
If now a weak rf field of frequency υ_{23} falls on the crystal, transitions are produced as before from level 2 to level 3 and also from level 3 to level 2. Because of the inverted population, however, there are now more transitions from level 3 to level 2 by emission of radiation of frequency υ_{23} than there are in the reverse direction. This is the basic amplification mechanism of a maser.

It is found experimentally that the thermal relaxation mechanisms are so potent at normal temperatures that strong enough pump power sources are generally not available for production of the inverted distribution. For this reason, and also because the inverted population is proportional to 1/T (see Equation 3), temperatures in the range of liquid helium (4.2°K and below) are most often used. However, masers have operated successfully at about 60°K.

In Figure 3 we show a schematic representation of a maser amplifier. The maser material is placed in a cavity which is resonant at the signal frequency. This is done in order to enhance the emission. It is also desirable to make the cavity simultaneously resonant at the pump frequency in order to be able to saturate the pump transition with a reasonable amount of pump power. The amount of pump power typically used is of the order of milliwatts. The output power is of the order of microwatts.

The quality of the maser material is describable in terms of conventional microwave terminology. In the inverted state the material is characterized by a negative Q which is defined as

$$-\frac{1}{Q_{maser}} = \frac{\text{Power emitted}}{2\pi\upsilon_{23} \times \text{Energy stored}} \qquad (4)$$



An engineer about to attach the ruby-filled microwave cavity to the wave guide and control assembly of a maser.

For a completely filled cavity this quantity can be computed to be

$$-\frac{1}{Q_{maser}} = \frac{16\pi^2(n_3 - n_2)\,|\mu_{23}|^2\,T_2}{h} \quad . \quad (5)$$

In Equation (5) μ_{23} is the quantum mechanical matrix element of the component of the dipole moment parallel to the rf field. The transition probability is proportional to the square of its magnitude. Actually μ_{23} depends strongly upon the orientation of the rf field with respect to the crystal axes and a proper orientation must be chosen to maximize the transition probability. T_2 is the so-called spin-spin relaxation time. It is related to the linewidth $\Delta\upsilon$ (as measured in a paramagnetic resonance experiment) by the equation $T_2 = 1/(\sqrt{3}\,\pi\Delta\upsilon)$.

Using an optimal orientation of the sample in the dc magnetic field and of the rf field, we compute for ruby with 0.05 per cent of Cr^{+++} ions, and for a frequency υ_{23} = 1300 mc, $Q_{maser} = -100$. The emission from the ruby will thus overcome cavity losses as long as the cavity's unloaded Q is larger than 100, and amplification will be obtainable.

It can be shown that the square root of the power gain g multiplied by the bandwidth is approximately a constant.

$$B\sqrt{g} \approx \frac{2\upsilon_{23}}{|Q_{maser}|} \quad . \quad (6)$$

With the above numbers we obtain, again at 1300 mc, a value of 26 mc. At higher signal frequencies this quantity is, of course, larger.

Noise in Masers

The inherent noise in a maser amplifier results from spontaneous transitions of ions from state 3 to state 2. This noise is present at all temperatures and it cannot be avoided. There are, of course, also contributions to the noise from the losses in the cavity. It may be shown that the noise figure has the form

$$\text{N.F.} = 1 + \frac{Q_{ext}}{Q_o}\frac{E_v(T_c)}{E_v(T_s)} + \frac{Q_{ext}}{Q_{maser}}\frac{E_v(T_x)}{E_v(T_s)}. \quad (7)$$

In Equation 7, Q_{ext} is the external Q and Q_o is the unloaded Q of the cavity. $E_v(T)$ stands for

$$h\upsilon/[\exp(h\upsilon/kT)-1] ,$$

T_c is the cavity temperature, T_s the source temperature, and T_x is the "negative temperature" existing between the inverted levels. It is common to define an effective noise temperature of the amplifier as that source temperature for which the N.F. just becomes 2. Consider an example: a ruby maser in liquid helium, signal frequency = 1300 mc, pump frequency = 11,160 mc, Q_o = 1000, Q_{maser} = − 200, Q_{ext} = 200. The noise temperature becomes 2.36°K. It can be seen that all reasonable combinations of Q_o, Q_{ext}, and Q_{maser} yield temperatures of a few degrees absolute. It can furthermore be shown that with vanishingly small cavity losses and complete inversion of the paramagnetic levels, and under the assumption of high gain, the effective noise temperature of the amplifier becomes

$$T_{eff} = \frac{h\upsilon_{23}}{k} \quad (8)$$

which represents some sort of an upper limit of the noise performance. For 1300 mc this upper limit lies somewhat under 0.1°K. From Equation (8) it is seen that T_{eff} increases with υ_{23}. Thus if we would build a maser at a wavelength of 5μ an upper limit for its noise performance would be represented by $T_{eff} \approx 3000$°K. From this figure it is quite obvious that it is completely impossible to use masers to detect jet airplanes or missiles by their black body radiation.

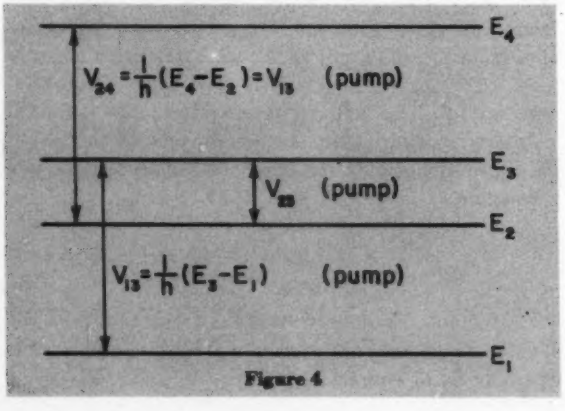


Figure 4

Some More Recent Developments

If the magnetic field makes an angle of 54° 44′ with the crystal c-axis in ruby, then the energy spectrum becomes symmetric, as shown schematically in Figure 4. Pumping between levels 1 and 3 also induces transitions between levels 2 and 4. This maser is sometimes referred to as a double pump maser. This double pump scheme is more effective in inverting the population of levels 3 and 2. Analogously to Equation (3) one obtains

$$n_3 - n_2 = \frac{N}{8}\frac{h}{kT}(\upsilon_{12} + \upsilon_{34}). \quad (9)$$

It is apparent that a negative temperature is always achieved between levels 3 and 2 and there is no such condition as previously that the pump frequency must be more than twice as large as the signal frequency. Such a double pump scheme would thus be especially important for masers at high frequencies where it is no longer possible to pump in excess of twice the signal frequency. However, the pump frequency still has to be larger than the signal frequency. Because of the larger amount of population inversion, such a double pump scheme is also desirable for maser operation at temperatures above 4.2°K. Actually the maser previously quoted as operating at 60°K used this double pump method.

It is also possible to place the maser crystal not into a microwave cavity but into a slow wave structure. In such a slow wave structure the interaction of the radiation with the paramagnetic centers is still sufficiently enhanced to observe maser action successfully. The advantage of a slow wave structure lies mainly in the fact that the bandwidth of the amplifier is no longer determined by the cavity but rather by the linewidth of the magnetic salt. This linewidth is of the order of 40 mc. Also, larger values of the square root gain-bandwidth product are possible. At S-band values in excess of 1000 mc have been reported.

Conclusion

The maser represents the lowest noise amplifier known at the present time. In the maser, for the first time in the history of the electronics industry, use is made of quantum transitions to achieve amplification. As a result, a new field which may properly be called quantum electronics has opened to the engineer and scientist. Many new and perhaps revolutionary devices should result from the further exploitation of quantum effects.

the PARAMETRIC AMPLIFIER

S. Cohen
Research Division

THE parametric amplifier has now been a part of the microwave art for a little over three years. Its advantages as a low-noise circuit element have been duly recognized and engineers are now being faced with the problem of using this negative impedance device as a stable amplifier. Let us briefly explain what a parametric amplifier is and describe some of its potential as well as its limitations.

A parametric amplifier is a method for obtaining voltage amplification by suitably coupling an energy storage element, whose value can be made to vary as a function of time, to a resonant circuit. If energy of the proper frequency and phase is supplied to the time-varying element, voltage amplification will occur in the tank circuit. This principle can be illustrated very simply by reference to Figures 1 and 2.

Let us assume that we physically move the condenser plates apart when the voltage v_t is a maximum (either positive or negative). Since the charge q remains constant, and C is decreased by the separation of the plates, the voltage across the tank circuit will increase. When the voltage v_t goes through zero, the plates are returned to their original position. The energy expended in pulling the plates apart appears in the electric field across the condenser plates and is transferred to the resonant circuit, producing voltage amplification at the resonant frequency and eventually oscillation. If the tank circuit is loaded properly the oscillation can be damped, and we have a parametric amplifier. This configuration requires pumping at exactly twice the resonant frequency and is sometimes called the degenerate case for reasons which will become clear immediately. Amplification is extremely sensitive to the phase of the pump with respect to the signal — in this case, a feature which may be troublesome, but which can be used to advantage in some applications, such as a Doppler radar receiver and in computer elements.

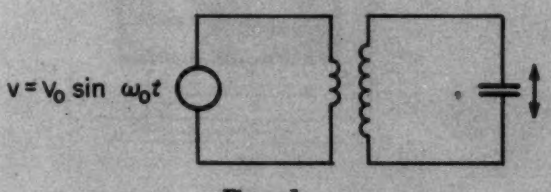


Figure 1

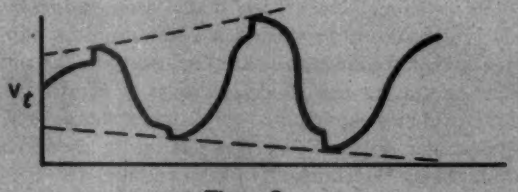
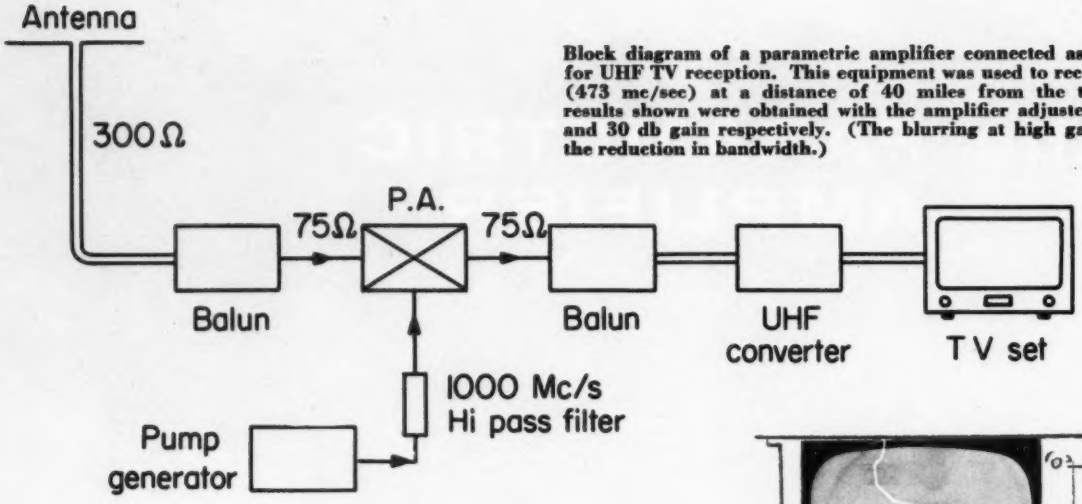
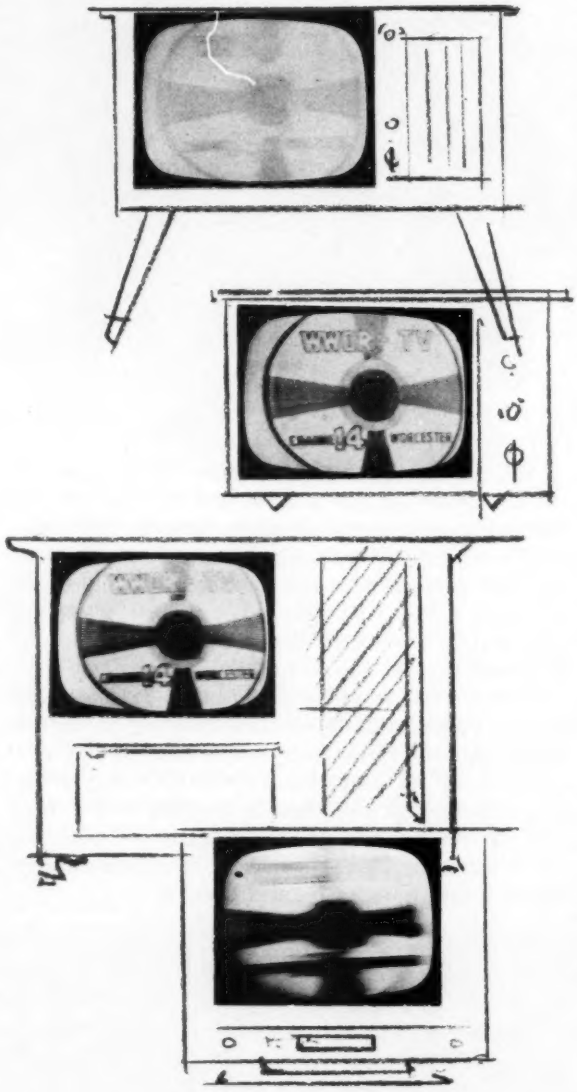


Figure 2



Block diagram of a parametric amplifier connected as a preamplifier for UHF TV reception. This equipment was used to receive Channel 14 (473 mc/sec) at a distance of 40 miles from the transmitter; the results shown were obtained with the amplifier adjusted for 5, 10, 20 and 30 db gain respectively. (The blurring at high gain is caused by the reduction in bandwidth.)

Any non-linear reactance that can be time-varied can be used as a parametric amplifier. This principle has been known for some time, but it was the introduction of the semiconductor diode that made parametric amplification at microwave frequencies possible. A variable capacitance diode is produced by adding impurity atoms to a semiconductor material so that a junction is formed, one side of which has an excess of electrons (n-type impurity) and the other has an excess of holes (p-type impurity). The process of diffusion of the two types of charge across the junction under normal equilibrium conditions (room temperature and zero external bias) results in restraining forces being set up at the junction that tend to keep the holes in the p-type region and the electrons in the n-type region. A transition region (often called the space-charge region) results, which is depleted of mobile charges, leaving only uncompensated, fixed charges. An electrostatic potential difference (potential barrier) now exists across this depletion region, giving rise to the zero-bias capacitance of the diode. If the potential drop across this region is varied by an externally applied voltage, the charge density in the depletion region will vary, resulting in a variation of the width of the space-charge region and a consequent variation of capacitance. The only limitation on the speed with which this capacitance can be varied is the "dielectric relaxation time" of the uncompensated charge, which is on the order of 10^{-13} seconds. Thus we see that a semiconductor diode is capable of producing a reactance that can be time-varied at high microwave frequencies. The equivalent circuit for this diode is shown in Figure 3. C_o is the zero bias capacitance (or fixed capacitance if dc reverse bias is used); C_3, the change in capacitance produced by the sinusoidal pump voltage, and R_b, the bulk resistance of the diode, including the contact resistances.



The phase restriction of the degenerate amplifier can be eliminated by using two (or more) resonant circuits coupled together by the variable reactance. If the variable reactance diode is sinusoidally pumped at a frequency $f_3 = f_1 + f_2$, where f_1 and f_2 are the resonant frequencies of the two tank circuits, the diode capacitance variation is given by $C = C_3 \sin(\omega_3 t + \phi_3)$. If we assume high "Q" tank circuits with f_1 and f_2 sufficiently far apart so that the two circuits are essentially independent, pumping the diode with sufficient amplitude will result in sustained oscillations at f_1 and f_2 in the respective tank circuits. If we couple a source and load into either of the tank circuits we have a non-degenerate, three-frequency, parametric amplifier. This circuit has been analyzed by Heffner and Wade,[1] and Chang and Bloom,[2] and will be briefly reviewed here.

Figure 4 shows the equivalent circuit for the non-degenerate amplifier, without a source and load. The fixed diode capacitance, C_o, has been included in the capacitances C_1 and C_2 of the resonant circuits. If we apply a voltage v_3, the pump voltage, at a frequency equal to the sum of the resonant frequencies of the two tank circuits, currents will be produced at the sum and difference frequencies due to the mixing action of the diode. Assuming high "Q" resonant circuits, only currents at the resonant frequencies, f_1 and f_2, will flow and produce voltage components across the capacitor C_3. The effective admittance presented to Tank 1 is a function of C_3 and the voltage across Tank 2. The voltage across Tank 2 is in turn a function of C_3 and the voltage across Tank 1. The admittance seen by Tank 1 at frequency f_1 turns out to be a negative real quantity, and is given by:

$$Y(\omega_1) = -\frac{\omega_1 \omega_2 C_3^2}{4G_2} = -G \qquad (1)$$

when the diode is pumped at exactly the sum of the two resonant frequencies.

The negative conductance is proportional to C_3^2, and inversely proportional to the conductance of the second tank circuit (referred to as the idler since it is not normally loaded). An equivalent circuit of the amplifier at resonance may now be drawn (Figure 5) from which the gain may be calculated: G_1 is the conductance of unloaded Tank 1, G_g and G_L are the source and load conductances. Since the impedance seen by the source at the signal frequency is negative, the input can never be matched, so that the only power gain that it is meaningful to talk about is the transducer gain, G_t, which is the ratio of P_{out} to the maximum available input power. For the circuit of Figure 5, this is:

$$G_t = \frac{P_{out}}{i_g^2/4G_g} = \frac{4G_g G_L}{(G_T - G)^2} \qquad (2)$$

where $G_T = G_1 + G_g + G_L$.

It is obvious from Equation (2) that the gain of the amplifier can have any value from 0 to infinity. The gain is dependent on the pumping amplitude, since this produces the capacitance variation C_3. This means that dC/dV for the diode must be as large as possible, or the diode must be biased in the region where the capacitance variation is a maximum (zero or small forward bias).

Equation (2) can be used to point out some basic limitations of parametric amplifiers operating at high gain levels. If we differentiate this expression, we obtain the gain sensitivity of the amplifier with respect to source impedance:

$$\frac{d(G_t)}{G_t} = -G_t^{1/2} \frac{d(G_g)}{G_g} \qquad (3)$$

For a transducer gain of 30 db, as an example, a 1 per cent change in source conductance results in a 32 per cent gain variation. An expression of the same form describes the dependence of gain on load conductance. Another factor limiting the usefulness of the parametric amplifier is the pump stability requirement. Since the negative conductance is produced by the time-varying capacitance of the diode, any variation of either pump amplitude or frequency can also produce large variations in gain, when the amplifier is operating at high gain levels. For the non-degenerate amplifier the variation of the pump amplitude

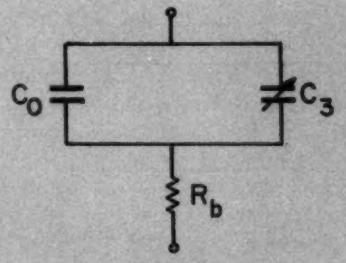


Figure 3

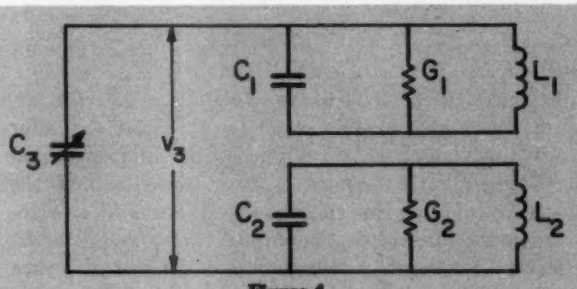
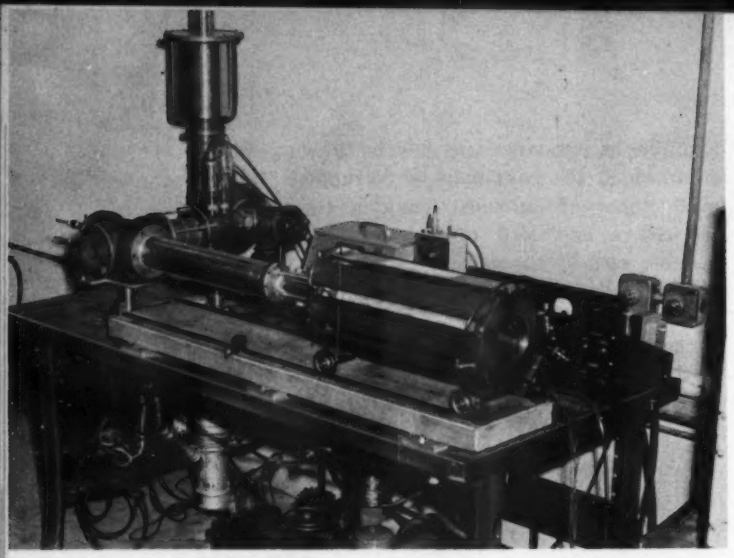


Figure 4



An experimental model of an electron-beam parametric amplifier in a demountable vacuum system. Here, parametric amplification keeps noise generated by the hot cathode from combining with the signal, as it does in the conventional traveling wave tube and klystron amplifiers.

with time is the more serious. This is not too difficult a problem to overcome except where the system requirements are severe.

There are two more items of vital interest to engineers concerned with using a parametric amplifier: (1) the gain-bandwidth of the device, and (2) the noise properties. The fractional gain-bandwidth product of the amplifier is given by the following:

$$\sqrt{G_t} \times \frac{\Delta f}{f_1} = \frac{2\omega_2 (G_g G_L)^{1/2}}{\omega_2 Q_1 G_T + \omega_1 Q_2 G} \qquad (4)$$

where Q_1 and Q_2 are the loaded "Q's" of the signal and idler tanks respectively, and Δf is the bandwidth at the half-power points. At high gain $G \cong G_T$, and with $G_1 << G_g + G_L$ the maximum gain-bandwidth occurs with $G_g = G_L \cong G/2$. The right side of Equation (4) then becomes:

$$\frac{f_2}{f_2 Q_1 + f_1 Q_2} \quad , \qquad (5)$$

and dividing by f_2 we get:

$$\sqrt{G_t} \times \frac{\Delta f}{f_1} = \frac{1}{Q_1 + \frac{f_1}{f_2} Q_2} \quad . \qquad (6)$$

To make the product a maximum, therefore, the circuit designer has two choices: (1) load the idler (reduce Q_2), and (2) increase the idler to signal frequency ratio (reduce f_1/f_2). Since loading the idler would require more pump power and also increase the idler noise, the obvious choice would be the latter. Fortunately, through one of Nature's rare coincidences, increasing the idler to signal frequency ratio also buys something the engineer is looking for, better noise performance, as will be seen in the next paragraph. A typical fractional gain-bandwidth product is about 0.01, achieved with $Q_1 \approx 40$, $Q_2 \approx 300$.

For a non-degenerate parametric amplifier, the noise figure is given by:[1, 2]

$$F = 1 + \frac{T}{T_o}\left(\frac{G_1}{G_g} + \frac{\omega_1}{\omega_2}\frac{G}{G_g}\right) \qquad (7)$$

where T_o is the standard noise temperature, 290°K, and T is the operating temperature. The terms in this expression are due to the noise contributions of the signal and idler circuits. G_1, the unloaded conductance of the signal tank circuit, is usually very small compared to the source conductance, G_g. The chief contribution to the amplifier noise is that due to noise generated in the idler circuit, which is given by the second term in Equation (7), $\frac{\omega_1}{\omega_2}\frac{G}{G_g}\frac{T}{T_o}$. As mentioned in the preceding paragraph, ω_1/ω_2 is made as small as possible for optimum noise and bandwidth. We can also reduce the idler noise contribution by making T as small as possible. If the effective noise temperature of the source, G_g, is very low (if the source, for example, is an antenna pointed skyward), and the idler circuit is loaded by coupling the idler to the source, this will effectively lower the operating temperature. However, if the diode bulk resistance constitutes the main idler loading, it is necessary to cool the diode itself.

The theoretical expression for the noise-figure, Equation (7), has been verified at the Research Division on two amplifiers.[3, 4] An excess noise figure (F-1) = 0.62, corresponding to an effective noise temperature [T_{eff} = (F-1)(290)] of 180°K and a noise figure of 2.1 db, was measured on a VHF coaxial-line amplifier with signal frequency 140 mc/sec and pump frequency of 615 mc/sec. A cavity-type amplifier with f_1 = 813 mc/sec and f_2 = 1996 mc/sec had a measured excess noise figure of 0.7, corresponding to an effective noise temperature of 203°K and a 2.3 db

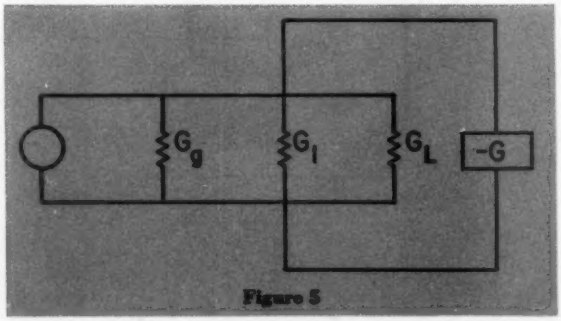


Figure 5

noise figure. These noise figures are the optimum that can be achieved, using either a microwave circulator or an output transformer to effectively open-circuit the load noise. The use of transformer coupling [5] to the output stage lowers the noise figure, but at the expense of the bandwidth.

Any article on parametric amplifiers is incomplete without mention of the traveling-wave amplifier and the amplifying up-converter. The band-width limitations imposed by the use of high Q signal and idler circuits can be overcome, without destroying the low noise of the device, by using a number of diodes in a traveling-wave structure. This device is capable of large bandwidth (30 per cent bandwidth has been reported) and unilateral amplification, and has become increasingly attractive. The amplifier is composed of a transmission line with variable reactance diodes acting as the shunt capacity. The diodes are sequentially pumped by means of a progressive wave traveling at the pump frequency, ω_3, having a phase constant, β_3. If a signal wave at frequency ω_1 can propagate down the structure with phase constant β_1 so that $\omega_1 + \omega_2 = \omega_3$ and $\beta_1 + \beta_2 = \beta_3$, where 1, 2, and 3 refer to signal, idler, and pump, respectively, then unilateral amplification at the signal and idler frequencies can occur.

General energy equations relating the powers at different frequencies in non-linear capacitors and inductors have been derived by Manley and Rowe.[6] If we use these relationships for the case where the variable reactance is pumped at a frequency equal to the difference of the two resonant frequencies, Figure 6, then the operation is stable, the amplifier input impedance is positive, and the maximum gain is equal to the ratio of the output to the input frequency. This amplifying up-converter is capable of wider bandwidth than the negative resistance amplifier but the input frequency for reasonable gain (10 db) is presently restricted to frequencies of the order of L-band due to limited availability of diodes capable of operation above X-band. The amplifier can be followed by a standard resistive type mixer to provide considerable improvement in noise figure over the normal combination of mixer and I.F. amplifier.

Traveling wave amplifiers have been constructed at frequencies well below the microwave region; amplifying up-converters seem by their nature to be destined for use at the low end of the microwave band. Single-cavity negative-resistance amplifiers, of the type discussed in the main part of this paper, suffer from the disadvantages of relatively small bandwidth and of high gain sensitivity. On the other hand, they seem to provide the most convenient way of achieving low noise amplification over a broad range of microwave frequencies.

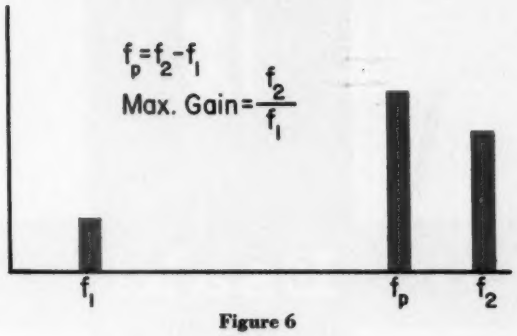


Figure 6

While presently available parametric amplifier diodes allow operation only up to about X-band frequencies, several companies are working intensively on the improvement of diode properties. The so-called cut-off frequency, f_c (defined as $1/2\pi R_b C_o$) is most important and shows what changes in diode parameters must be achieved to permit operation at still higher frequencies. As higher cut-off frequencies are obtained, the "Q" of the diodes at any given frequency will increase. Performance in general will be improved and especially with respect to noise properties and pump power requirements.

[1] H. Heffner and G. Wade, "Gain, Bandwidth, and Noise Characteristics of the Variable-Parameter Amplifier," J. Appl. Phys. *29*, 1321, Sept. 1958.

[2] S. Bloom and K. K. N. Chang, "Theory of Parametric Amplification Using Nonlinear Reactances," R.C.A. Review, Dec. 1957.

[3] S. Cohen, "A VHF Coaxial Line Parametric Amplifier," Raytheon Company, Research Division Technical Memorandum T-164.

[4] C. Freed, "The Theory of Noise Measurements on Cavity-Diode Type Parametric Amplifiers and Experimental Verification of Noise and Gain-Bandwidth Theory," Raytheon Company, Research Division Technical Memorandum T-149.

[5] H. A. Haus, "Theory of Noise Measurements on Parametric Solid-State Amplifier of Transmission Cavity Type," Raytheon Company Technical Memoranda T-138 and T-138-A.

[6] J. M. Manley and H. E. Rowe, "Some General Properties of Non-Linear Elements," Proc. IRE, *44*, 904, July, 1956.

A parametric amplifier

the Esaki "Tunnel Diode"

R. Pucel
Research Division

THE Esaki p-n junction diode is the most recent addition to the class of negative resistance solid state devices used in low-noise microwave amplifiers. Named after its inventor,[1] this device is also known as the "tunnel diode" because its operation is based on the quantum mechanical "tunnel effect." Because of the Esaki diode's low noise properties (still unrealized), its potential application as a low-level (milliwatt) microwave amplifier and oscillator, and its probable use as a high-speed bistable computer switch, it is being intensively investigated in the Research Division, as well as in other areas of Raytheon and at other industrial firms.*

As is probably well known to most readers, a poorly conducting piece of chemically-pure semiconductor, such as germanium or silicon, can be converted to a reasonably good conductor by the insertion of certain known "impurity" or "doping" atoms which substitute for some of the parent crystal atoms. If now instead of one doping agent a proper combination of two "unlike" dopents is used, one dopent say for each half of the parent crystal, the material can be made to conduct very heavily in one direction and very poorly in the other, a property basic to a diode. A typical diode characteristic achieved with semiconductors is shown in Figure 1.

The cross-sectional plane separating the two regions of opposite type impurities is known as the p-n junction. A consequence of the difference in impurities flanking the junction is that a narrow electrically charged dipole region (on the order of 10^{-5} cm) is formed, giving rise to a small potential hill or barrier at the junction. This barrier prevents a net current from flowing across the junction at thermal equilibrium (that is, with no bias voltage applied). The width of this junction depends, among other things, on how highly doped the two halves are. The higher the doping, the narrower the region will be.

In a conventional p-n diode, the number of charged carriers which surmount the barrier can be varied by changing the height of the barrier with an applied

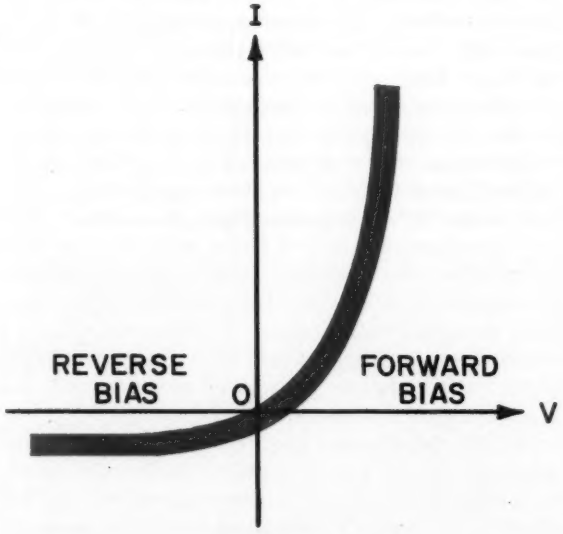


Figure 1

* It may turn out eventually that the most useful application of this diode will be in the field of ultra-high speed computer switches. Experiments have been reported which show that these diodes, used as two-terminal bistable switches, can be operated at switching rates corresponding to several hundred megapulses per second, many times faster than existing computer switches. These rates can undoubtedly be increased. Subharmonic parametric oscillators operating in approximately the same frequency range are also being considered as computer switches. However, they will involve more circuitry and a very stable pump.

bias. When the barrier is lowered, a large "forward" current is made to flow. When the polarity of the bias is reversed so that the barrier is raised, a smaller reverse current is forced through the junction, hence the positive resistance diode characteristic mentioned previously. Because the carriers flow across the junction by a combination of drift and diffusion, this diode is often called a diffusion diode.

If the concentration of dopents is increased sufficiently, an additional mechanism of charge flow is introduced. It is then not only possible for some carriers to flow over the barrier (which now is of the order of 10^{-6} cm wide) but also *through* it. Classically, of course, this is an impossibility; however, since charge carriers in a crystal obey the laws of quantum mechanics, their behavior is not restricted by classical laws. In the language of quantum theory, there is a small but non-zero probability for electrons to make a "transition" from one energy band, the valence band on the p-side of the junction, across a forbidden gap, to another energy band, the conduction band, on the n-side of the junction and vice versa, provided the dopent concentrations are high enough to cause these bands to overlap in energy.

Since tunneling can occur in either direction across the junction, the net tunnel current consists of the difference of two oppositely-directed components, which, of course, cancel at thermal equilibrium. One component or the other is made to dominate when a bias voltage is applied. In the reverse biased direction the tunnel current increases rapidly with bias, giving rise to the Zener breakdown phenomenon. With applied bias in the forward direction, however, it turns out that the net tunnel current at first increases with bias voltage, and then decreases when the bias voltage exceeds a certain value, because the number of electrons available for tunneling is reduced. The composite I-V characteristic, comprising both the conventional diffusion current and the tunnel current, is shown in Figure 2.

We are interested only in the negative resistance region produced by tunneling, that is, where the slope dI/dV is negative. This region extends typically over the voltage range from a few tenths of a volt to about half a volt, depending on the semiconductor and concentration of dopents used. The current range depends on the latter factors, and in addition, on the junction cross-sectional area. In a well-designed Esaki diode, the diffusion current component is usually negligible in the negative resistance region.

If the quiescent operating point is adjusted to fall in this region, for small signal excursions about this point the p-n junction can be represented as a circuit element by a negative conductance equal to the slope of the curve, dI/dV, in parallel with the usual capacity, common to all p-n junctions. To this must be added in series an element representing the (low) series resistance of the semiconductor bulk external to the junction and of the electrical contacts to the former. Combining these elements, we obtain the small signal ac equivalent circuit shown in Figure 3.

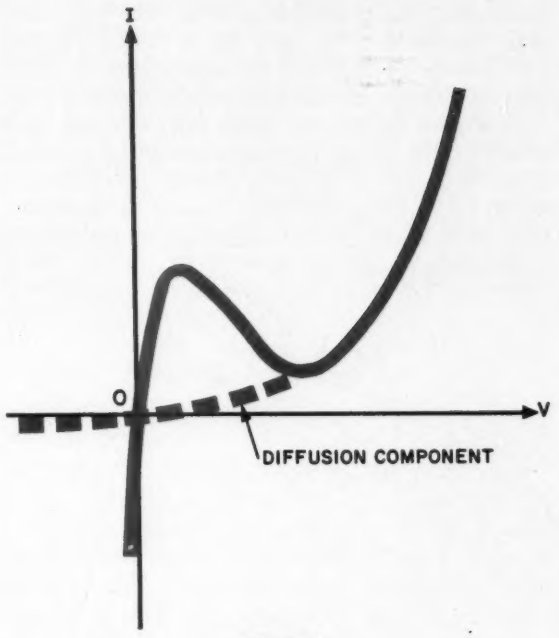
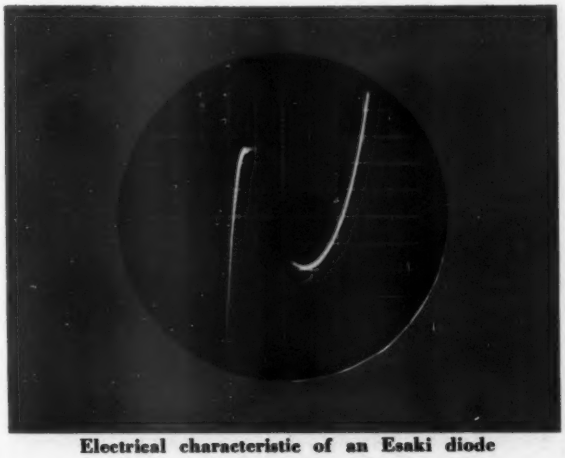


Figure 2

Typical values of negative resistance R (equal to G^{-1}) range from a few tenths to several hundred ohms. Capacity (C) values are more restricted and range typically from about 10 to 1000 μμf. The bulk resistance R_b varies the least and is usually of the order of 0.1 ohm. All of these quantities depend, of course, on the choice of materials and on geometrical



Electrical characteristic of an Esaki diode clearly shows the negative resistance region.

factors such as cross-sectional area. The bulk resistance, because it is so small due to the high doping, can usually be neglected in calculations of circuit behavior, except for the following situation.

If it were not for this series resistance, the diode would exhibit a negative resistance at its terminals at all frequencies. However, because R_b is non-zero, it will cause the negative resistance of the series combination to vanish at some critical or cutoff frequency f_c given, approximately, by

$$f_c = \frac{1}{2\pi C}\sqrt{\frac{G}{R_b}} = \frac{1}{2\pi C\sqrt{R\,R_b}} \quad . \quad (1)$$

Above this frequency, the Esaki diode cannot be made to amplify or oscillate.

Thus, even though the tunneling mechanism imposes no essential frequency limitation, the external losses do in the presence of the unavoidable junction capacity. From Equation (1) it is clear that the lower these losses are, the higher the frequency at which the diode can be made to operate. The same is true for the capacity. It is equally true that the greater the tunnel effect (that is, the larger the magnitude of the slope dI/dV, or G, per unit capacity) the higher the upper frequency limit. It is the task of the device engineer to choose the appropriate materials for realization of a high G/C ratio. We shall find later that this goal is also desirable for other reasons. Diodes have already been reported which oscillate at microwave frequencies (4 kmc).[2] We shall assume in the discussions which follow that the frequency of operation is well below the cutoff, so that the series resistance may be neglected.

Since by definition a negative resistance is a source of power (as contrasted to a positive resistance which is an absorber), one may use the former as the building block of an amplifier by simply connecting a signal source and a load in parallel with it, as shown in Figure 4. The ideal current generator I_s and conductance G_s constitute the signal source and G_L represents the load. The inductance L resonates with the diode capacity C at some frequency

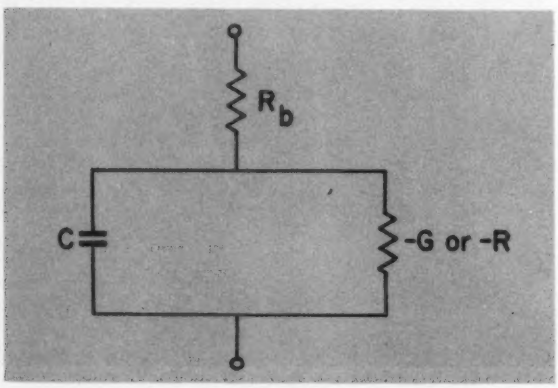


Figure 3

$$f_o = (2\pi\sqrt{LC})^{-1}.$$

The power amplification arises as follows: The signal generator develops a voltage across the negative conductance and the load. The former, because it is negative, passes a current at the frequency of the voltage but 180° out of phase with it; that is, it forces power back into the source and, of course, into the load. If conditions are right, it will send much more power into the load than is available from the signal source; in other words, it produces power gain. Of course, this additional power ultimately comes from the dc bias supply.

Because of the tuned circuit, the power gain will be maximum at the resonant frequency, falling off on either side in a fashion typical of such a circuit. Since the amplifier is of the negative resistance type, one cannot "match" it to the source and load; therefore, the conventional definition of available power gain is inappropriate. Instead, we use as our measure the so-called transducer gain $G_t(\omega)$ which is defined as the ratio of the power delivered to the load divided by the maximum power that could be absorbed from the source if it were matched; namely, $I_s^2/4G_s$. At resonance this gain is equal to

$$G_t(\omega_o) = \frac{4G_sG_L}{(G_L+G_s-G)^2} \quad (2)$$

and the bandwidth, or frequency difference between half-power points, is

$$B = \frac{G_s+G_L-G}{2\pi C} \quad . \quad (3)$$

Inspection of these two equations illustrates the behavior typical of all amplifiers of the negative resistance type; as the total loading $G_s + G_L$ approaches the negative conductance in magnitude, the gain at resonance increases to infinity and the bandwidth reduces to zero. In essence, the "effective Q" of the loaded tuned circuit becomes arbitrarily large. At the compensation point, $G_s+G_L = G$, when the net conductance vanishes, the gain is infinite, and an oscillator results.

As is typical for many types of amplifiers, whether of the negative resistance variety or not, resonance gain is traded for bandwidth. In other words, the merit of the amplifier can be summarized by a constant gain-bandwith product which characterizes the amplifying device alone. In our case, the maximum value of this

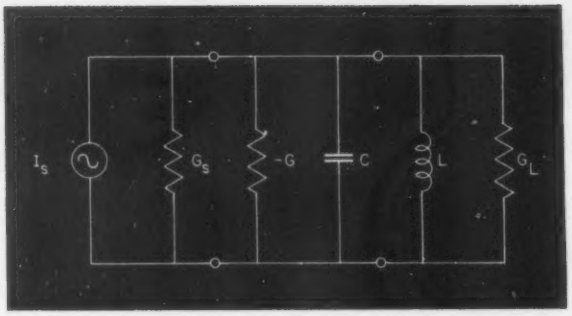


Figure 4

product (at high gain) occurs when the loading is shared equally between the source and load, that is, when $G_s = G_L \approx G/2$. It is given by

$$\left(\sqrt{G_t(\omega_o)}\ B\right)_{max} = \frac{G}{2\pi C} \quad , \qquad (4)$$

which depends only on the parameters of the diode. Equation (4) is a useful quantity for the device engineer and again clearly illustrates the need for a high G/C ratio. Gain-bandwidth products of the order of 1 kmc have been reported.[3] Symmetric loading, though optimum for gain-bandwidth considerations, is not optimum for best noise performance, as will be discussed below.

There are two sources of noise in the Esaki diode, a thermal noise generated in the series bulk resistance R_b and proportional to it, and a shot noise produced in the junction itself. The shot noise contribution is by far the dominant one because of the small value of R_b. It arises from the fact that the electrons which cross the junction and constitute the bias current do so at random time intervals; therefore, fluctuations are superimposed on the average junction bias current. This same effect occurs in other devices such as vacuum diodes. These fluctuations, of course, are independent of the direction of charge flow. It turns out that, from a circuit standpoint, these fluctuations can be represented by a noise current generator in parallel with the negative conductance and shunt capacity. Its rms amplitude is nearly proportional to the sum of the magnitudes of both tunnel current components, which sum in turn can be approximated, for a well-designed diode, by the net bias current I.

The noise performance of an amplifier can be characterized by its noise figure F which is simply the ratio of the noise power delivered to the load by all sources, exclusive of noise contributed by the load itself (that is, the contribution of the noise of the signal source plus that of the amplifier) divided by the noise contribution of the signal source alone, assumed to be at room temperature, 300°K. (In practice, one cannot ignore the noise contributed by the load since it is also amplified, even though according to the standard definition it should not be included. At frequencies where microwave circulators are available, however, this load contribution can be eliminated.) It turns out that for the Esaki amplifier (and other negative resistance amplifiers, such as the parametric amplifier), the minimum noise figure occurs for highly unsymmetric loading, specifically, when nearly all of the loading is due to the source, $G_s \approx G >> G_L$. This minimum figure is given simply by

$$F_{min} \approx 1 + \frac{20\ I}{G} \qquad (5)$$

at room temperature, where I is approximately the operating bias current expressed in amperes, and G is the magnitude of the negative conductance in mhos.

Though this minimum value is subject to some optimization by appropriate choice of materials and proper design of the diode, it is not as strongly dependent on these factors as are the cutoff frequency and gain-bandwidth product. Obviously, the device engineer should strive for a high conductance per unit current. The best noise figure reported to date is slightly over 2 db, an effective noise temperature of 200°K.[4] However, more typical of the numbers achieved are noise figures of the order of 4 to 6 db and these are not particularly impressive. It will be difficult to attain the best noise figures now available with parametric amplifiers.

Inventions

Congratulations to the following persons for their valuable contributions to Raytheon's portfolio of inventions:

Name	*Field of Invention*
Colin Bowness	Ferrite Switch
Alan W. Ede	R. F. Window
Paul X. Flaherty	Wetting of Silicon
Mario S. Floridia	Magnetron Beam Switching Tube
B. B. Frustajer	Masking of Transistors
Robert A. Fryklund	Radar Display
J. S. Gillette	Semiconductor Junctions
Irving Goldstein	Low Power Klystron
Thomas Gross	Radar Doppler Reference
Harold M. Hart	Display System
Rudolf Hergenrother	Permanent Periodic Magnet
Rudolf Hergenrother	Backward Wave Oscillator
James Lapointe	FM Oscillator
Jerome M. Lavine	Unipolar-Bipolar Field Effect Transistor
Jerome M. Lavine	Method of Carrier Injection
Max Michelson	Ferrite Circulator
Charles Philips, Howard Scharfman	Temperature Compensation
Wilhelm Rindner	Bonding Technique
Daniel Schiff	Bonding Method
Allen H. Smith	Silicon Carbide Crystals
Theodore C. Taylor	Semiconductor Structure
W. H. Thies, Jr.	Doppler Detector
Edwin E. Turner	Stabilized Radar Displays

Conclusion

The Esaki diode is very well suited for negative resistance amplifier and oscillator applications. It is quite simple and inexpensive to fabricate requiring, basically, only refinements in present techniques used in the manufacture of ordinary semiconductor diodes.

Unlike the parametric amplifier, it is free of the requirement for an rf pump source and the associated difficulties, such as gain variations arising from pump fluctuations. However, like the parametric amplifier (and the maser) it is still plagued with the effects of load and source impedance fluctuations on gain stability. The effects of load variations can be overcome, however, at frequencies where circulators are available. The diode's inherent frequency limitations will probably permit applications at frequencies as high as for parametric amplifiers and masers, and possibly higher, because a high frequency pump is not needed, a fact which will become important at millimeter frequencies.

Counteracting this latter advantage is a serious practical difficulty, namely, that these diodes, because of their relatively large shunt capacity, present abnormally low impedance levels at microwave frequencies. These low levels lead to difficult matching problems using conventional equipment and also prevent the design engineer from using, for example, microwave diode cartridges because of their inherent inductance. In fact, it appears that no lead lengths will be permissible at all. Diodes will probably be mounted directly in (low) impedance lines and waveguides. This technique, in which a diode is soldered directly into a microstrip line, has met with success in various laboratories.

The dynamic signal range is definitely restricted to lower level operation than is the case for parametric amplifiers. In preamplifiers, this should not be a serious disadvantage. However, the low output as an oscillator (below a milliwatt) is somewhat disappointing.

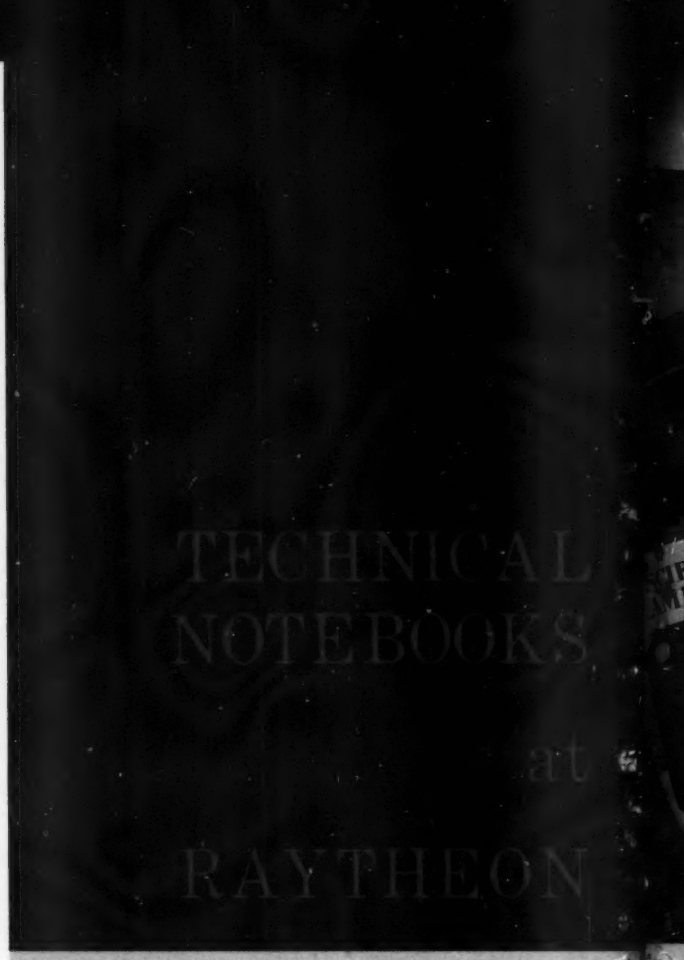
Where low noise figures are concerned, the Esaki amplifier cannot be considered as a too serious competitor of the best parametric amplifiers, and certainly not of masers. However, in applications where low-noise requirements are not sufficiently stringent to warrant the added complexity of the latter types, but still cannot be fulfilled by conventional amplifiers, the Esaki amplifier is admirably suited.

[1] L. Esaki, "New Phenomenon in Narrow p-n Junctions," Physical Review (L), Volume 109, page 603.

[2] 1959 Solid State Device Research Conference, Cornell University.

[3] H. S. Sommer, Jr., "Tunnel Diodes as High Frequency Devices," Proc. IRE, page 1201, July, 1959.

[4] 1959 Solid State Device Research Conference, Cornell University.



January 1, 1960 will mark the initiation of a new Technical Notebook program at Raytheon. Designed to benefit both the engineers and scientists as well as the Company, the program was motivated by Raytheon's explosive growth which placed undue strain on a system designed for a smaller concern and a slower tempo of technical activity.

In science the work of any one man rests on the foundation laid by his predecessors. Progress in science and technology arises out of past experience painstakingly absorbed and pondered. The natural result has been the great emphasis in every branch of science on "writing it down," recording every bit of experimental data and significant calculations in a permanent consolidated form. In industry, records of technical and scientific work have the same significance: to document technical progress, to ensure the continuity of effort and to avoid duplication. In addition good records serve to aid in the protection of the Company's interests. The primary users of these records therefore, are first and foremost the engineers of the Company, who can do an immeasurably better job if they maintain good records of personal activity and have ready access to similar records of others. Also of great importance is the use of this information

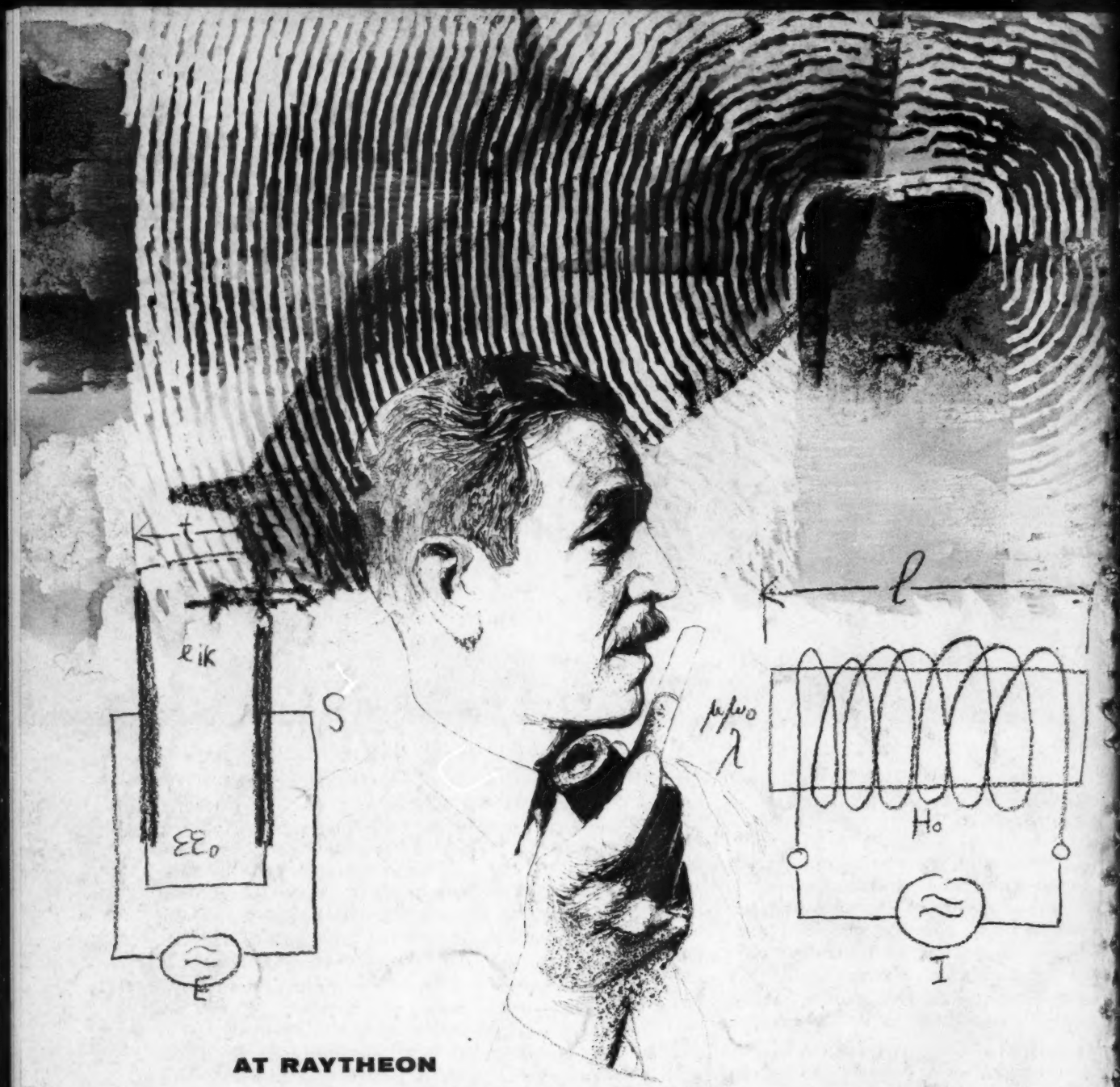
by Company management in the determination and protection of Company interests in proprietary and patentable areas.

The heart of this area of scientific documentation is the engineer's personal record of daily work — the Technical Notebook. This record, if properly kept, can unfold the complete course of each technical activity and present the full picture of the Company's technical progress. It provides program continuity even though people change; it is invaluable as legal evidence when defending the Company's interests.

Raytheon's new program centers around an entirely new notebook — one designed with the help of a group representing the Company's research and engineering areas. The importance Raytheon management accords the notebook is reflected in its general appearance and format. Considerable thought went into the design of pages resulting in an attractive layout which at the same time makes no compromise with usefulness. To preserve the legal integrity the books are hard bound and permanently sewed. Even so, special construction provides books which lie essentially flat when open.

All Technical Notebooks are the property of Raytheon. They will be issued to all who are performing engineering and scientific studies and tasks. Thus the holder is in essence a custodian, and is responsible for maintaining and safeguarding it. To provide the greatest benefit certain basic rules must be followed. First is the necessity for treating the book as a daily work diary. An entry should be made for every working day. The entries should, of course, be dated and written so as to provide the best possible engineering and scientific record. If you have properly kept your notebook, at a later date someone else should be able to accurately reconstruct what you did, what you thought, what you saw, who was present, to whom you spoke, what you discussed and where and when all of this occurred. This basic procedure and outlook hold true whether the book is kept by a physicist doing basic research or by a production engineer solving problems of production layout. The holder should make every effort to index the contents well in the space provided so that the book will be useful to others. This will be of help to the holder as well, since the program provides for his notebook (or copies) to be available to him as long as he is at Raytheon.

To be successful this program will require the fullest cooperation of everyone. The benefits accruing, both to Raytheon's engineering and scientific fraternity and to the Company, will make such cooperation well worthwhile.



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Among the distinguished scientists at Raytheon who are expanding basic knowledge of sonics is Dr. Theodor F. Hueter of the Government Equipment Division's Sonar Department.

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